SYSTEM AND METHOD FOR AN IF-SAMPLING TRANSCEIVER

TECHNICAL FIELD OF THE INVENTION

The present invention is generally related to radio transceivers and, more particularly, to a system and method for an intermediate frequency transceiver.

BACKGROUND OF THE INVENTION

Wireless technologies have seen significant improvement over the past several years. All sorts of communications devices are now seen as potential candidates for the installation of a wireless communications device. From telephones, to computers, to personal digital assistants, the list of wireless devices grows everyday. As merely an example, Bluetooth wireless local area networks purport to enable the installation of wireless devices into everything from jewelry to major appliances.

The Institute of Electrical and Electronics Engineers (IEEE) has developed new wireless ethernet standards under 802.11, which includes IEEE 802.11a, some of which have begun gaining acceptance in the industry. Even further, the European Telecommunications Standards Institute (ETSI) has developed a high performance radio local area network (HiperLAN). HiperLAN has an embodiment called HiperLAN/2, which is seen as being in direct competition for the widespread acceptance of the IEEE 802.11a standard. Both transmit in the 5 GHz unlicensed national information infrastructure (UNII) frequency range, and have data rates of about 54Mbps, and share other similarities at the physical layer. For example, both standards use orthogonal frequency division multiplexing. This means that the design of the radio architecture in both systems can have certain commonalities.

These commonalities are fortunate, because as transmission frequencies and data transfer rates rise, the complexity of the underlying radio architecture necessarily rises. The 802.11a and HiperLAN standards require especially complex solutions for the standards to be met. These complex solutions increase cost, which in turn increases the time needed to gain widespread acceptance in the industry.

Regular transceiver architectures employing I/Q down-converters cause problems in orthogonal frequency division multiplexing (OFDM) because they produce I/Q imbalance and DC offset, whereas this particular OFDM requires these values to be extremely low in order to obtain the specified signal to noise ratio (SNR). Intermediate frequency (IF) sampling architectures solve these problems, but introduce problems of their own, such as higher conversion speed causing increased analog-digital converter power consumption, and higher required selectivity to avoid both aliasing and image leakage. The higher selectivity requirement usually leads to using two intermediate frequency-surface acoustic wave (SAW) filters. These filters however, lead to increased noise figure and higher cost. Thus, an unaddressed need exists in the industry for an IF sampling architecture that obviates these problems.

SUMMARY OF THE INVENTION

The present invention provides a receiver portion of a transceiver that receives data via radio frequency transmission. The receiver comprises an IF sampling architecture, a quantizer, a baseband converter and a filter. The IF sampling architecture, wherein the IF sampling architecture receives the input signal, passes the input signal through a first filter, which is characterized by steep selectivity and narrow bandpass, converts the filtered signal to an IF signal and passes that signal through a second filter

that has a bandpass characteristic, but without the steep selectivity characterizing the first filter. Then, a quantizer digitizes the filtered IF signal, and a baseband converter converts the digitized signal to a baseband signal. Finally, a third filter filters the adjacent channel harmonics from the baseband signal to produce a data signal.

The invention further includes a method for receiving a radio signal. The method comprises the steps of receiving an input signal. Then, the input signal is filtered by in a first filter having a response characterized by a steep selectivity and a narrow bandpass. The filtered signal is then modulated to produce an in-phase and a quadrature phase IF signal at an intermediate frequency. Next, the IF in-phase and quadrature phase signals are filtered in a channel selection filter and the results of the channel selection filter are added together. Finally, the method comprises digitizing the sum and modulating the digitized sum to obtain a baseband in-phase data signal and a baseband quadrature phase data signal.

These and other features and advantages of the present invention will become apparent from the following description, drawings and claims.

BRIEF DESCRIPTION OF THE DRAWINGS

The invention can be better understood with reference to the following drawings. The components in the drawings are not necessarily to scale, emphasis instead being placed upon clearly illustrating the principles of the present invention. Moreover, in the drawings, like reference numerals designate corresponding parts throughout the several views.

FIG 1 shows a block diagram of a radio in which the transceiver of present invention resides.

FIG 2 shows a graph illustrating the signal strength requirements of the IEEE 802.11a and HiperLAN standards.

FIG 3 shows a graph illustrating the transfer function of an intermediate frequency SAW filter used in radio design.

FIG 4 shows a graph illustrating the signal strength requirements of the IEEE 802.11a and HiperLAN standards after taking into account the application of the intermediate frequency SAW filters.

FIG 5 shows a schematic diagram of the radio receiver of the present invention.

FIG 6 shows a graph illustrating the transfer function of a complex domain filter used in conjunction with the present invention.

FIG 7 shows a schematic diagram of the present invention with reduced complexity.

FIG 8 shows a schematic diagram of the radio of the present invention including a transmitter that operates in accord with the receiver of previous figures.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

The preferred embodiments of the invention now will be described more fully hereinafter with reference to the accompanying drawings, in which preferred embodiments of the invention are shown. The invention may, however, be embodied in many different forms and should not be construed as limited to the embodiments set forth herein; rather, these embodiments are provided so that this disclosure will be thorough and complete, and will fully convey the scope of the invention to those skilled in the art. Furthermore, all "examples" given herein are intended to be non-limiting.

The present invention is particularly suited for use in conjunction with either the Institute of Electrical and Electronics Engineers (IEEE) 802.11a standard for wireless communications or the European Telecommunications Standards Institute (ETSI) High performance radio local area network (HiperLAN)/2 standard. However, the present invention is not limited to use with these standards and can be modified to be suitable for other uses, as will be understood by those skilled in the art in view of the present disclosure. Both standards are actually competing for acceptance by the industry and consumers. As such, an cost effective solution to the radio architecture needs of both standards is the most efficient solution to the problem.

Both of the systems operate on radio frequency bands in the range of 5.1 GHz to 5.9 GHz. Each band inside this range has eight separate channels, with each channel slightly overlapping the channels on either side. These channels on either side are referred to as "adjacent channels." The HiperLAN/2 standard has more stringent guidelines for the rejection of the adjacent channels than does the IEEE 802.11a standard. Thus, when the present invention was designed, it was designed to meet the higher signal to noise ratios (SNR) required by the HiperLAN/2 standard, which, of course, enables it to meet the less stringent SNR requirements of the IEEE 802.11a standard.

Each channel is about 17 MHz wide, has a frequency spacing of about 20 MHz and is made up of 52 narrow band carriers which are about 300 kHz wide. Each of the narrow band carriers operate on a direct sequence spread spectrum protocol. These narrow band carriers use a coded orthogonal frequency division multiplexing scheme (COFDM) to encode the data that is being sent. All of the narrow band carriers are used, such that the system can send a number of data signals in parallel. As one skilled in the

art will recognize, the parallel transmission of data can occur much faster than transferring the data in sequence.

Common architectures used for these wireless systems employ an I/Q down converter. However, this causes problems with respect to OFDM radios. In these radios, the SNR is required to be on the order of about 30 dB or greater after analog to digital (AD) conversion. This requirement, because of the high frequency signals involved, requires that both the I/Q imbalance and DC-offset values in the radio be extremely low. The only way to solve this problem is through the use of complex compensation algorithms. These complex compensation algorithms in turn increase the cost of the radio.

In accordance with the present invention, generally IF sampling architecture digitizes the intermediate frequency signal with an AD converter clocked at a rate four times higher than its center frequency. The architecture of the present invention solves both of the aforementioned problems because DC offset is non-existent in the approach taken by the present invention, and the required I/Q imbalance is obtained by digitally converting the IF signal to baseband.

Sampling at four times the rate of the center frequency, the digital conversion to baseband is very cost effective to build. However, clocking the AD converter at this rate usually leads to higher power consumption. Moreover, higher selectivity is employed in the filtering steps to avoid both aliasing and image-leaking. The higher selectivity requirement most often leads to the use of two intermediate frequency surface acoustical wave (SAW) filters. These types of systems usually also require a “super-heterodyne” front end to convert the incoming signal at the 5-6 GHz range to a first intermediate

frequency of about 1.5 GHz. However, the use of the extra SAW filter usually provides more selectivity than is necessary for the reception of the signal.

Using the IF sampling architecture of the present invention, the first intermediate frequency signal is down converted to a second intermediate frequency. The second intermediate frequency is a design selection. As with all selections, there is a tradeoff. Here, the tradeoff is between performance and required selectivity. As was mentioned before, in accordance with this embodiment of the present invention, the sampling rate preferably is 4 times the second IF in order to provide high quality digital down conversion. Increasing the second intermediate frequency will necessitate a higher sampling rate, which might increase power consumption. Alternatively, lowering the second intermediate frequency might bring the image and alias frequencies closer to the wanted signal, thus increasing the required selectivity of the filtering devices. The present invention balances these tradeoffs in order to optimize performance and power consumption. One solution posed would be to under-sample the signal at a higher second intermediate frequency. However, this raises the problem of noise folding, wherein the noise figure increases. The manner in which these problems are handled to achieve an optimum solution will now be described with respect to example embodiments of the present invention.

Referring now to FIG 1, shown is a block diagram of the radio architecture in which the present invention resides. Wireless networks communicate from various endpoints to another endpoint that is typically hardwired to a network. Each of these wireless endpoints has a radio 101 installed into the endpoint to enable it to communicate with any of the other devices on the network. The radio communicates with other radios

on the network via radio frequency signals received and transmitted over an antenna 102. In this embodiment, the radio is designed for the industry standards IEEE 802.11a and HiperLAN/2, which both operate in about the 5.1-5.9 GHz range. In order to operate more efficiently, an incoming radio signal is first down converted to an intermediate frequency with a super-heterodyne front end 103. If the incoming RF signal is not down converted, the system would require a large amount of power to operate because of the relatively high clocking frequency that would be required. Again, it should be understood by one skilled in the art that the operating frequencies described herein are merely examples of possible operating environments in which the present invention can be used, and that there are other environments in which the present invention can be used without departing from the teachings contained herein.

After down converting, the present invention uses a intermediate frequency sampling architecture 104. This is intended to filter out the alternate, adjacent channels, and the alias and image channels that result from the modulation of the signal. Optionally, an I/Q sampling 105 architecture can be added. This will provide some extra selectivity by slightly amplifying the wanted signal while suppressing some of the unwanted signals. The signal is also digitized at stage 105 prior to baseband conversion and a final filtering 106 of the unwanted harmonics, resulting in a received signal (Rx) 107.

On the transmission side, what is close to a mirror image of the receiver exists. The transmitted signal (Tx) 108 is first filtered and modulated from the baseband signal into a second intermediate frequency and then converted from a digital to an analog signal by circuit 109. The signal is then modulated to a first intermediate frequency by

modulation circuitry 110. The signal is then amplified and filtered at the first intermediate frequency by amplification/filtering circuitry 111. Finally, the signal is modulated up to transmission frequency, filtered and amplified to transmission power by circuit 112. The signal is then transmitted to another endpoint enabled with a similar radio to communicate via radio frequency signaling. The focus of the present invention is on the receiver portion 113 of the transceiver.

Referring now to FIG 2, shown is a representation of the signal strength at the frequencies surrounding the wanted signal. As can be seen, the wanted signal 200 is at the local oscillator frequency (LO) plus 15 MHz. This frequency was chosen according to the frequency response of an intermediate SAW filter. Generally, the SAW filters used for this implementation show a flat frequency response for signals of about 20-30 MHz to either side of the center frequency. One skilled in the art should understand, however, that in accord with the previous paragraphs, this frequency is chosen according to design preferences, and thus could vary widely according to preference and use. The adjacent channels 201, 202 signal strength is shown at +21 dB because according to the standard, the adjacent channels 201, 202 can be up to 21 dB stronger than the strength of the wanted signal 200. Since the required SNR for the wanted signal 200 is 30 dB, this means that there must be at least 51 dB rejection at the adjacent channels 201, 202. Similarly, the alternate channels 203, 204 are shown at +40 dB, and therefore the required frequency rejection for the alternate channels 203, 204 is 70 dB. With respect to the image and alias responses shown, assuming half of the alternate channel power passes through to the image and alias channels, that equates to 3 dB loss, which means the image and alias channels are at +37 dB relative to the wanted signal. In turn, the system

requires that the SNR of the wanted signal at least 30 dB, so the required selectivity at these channels is 67 dB.

By closely observing the transfer function in FIG 3 of the intermediate frequency SAW filter applied at the first stage, the rejection obtained at relevant frequencies can be ascertained. Assuming an appropriate center frequency (in the 1.5 GHz range for this embodiment), the lower alternate channel sees 56 dBs of attenuation, the lower image channel sees 44 dBs of attenuation, the lower adjacent channel sees 15.7 dBs of attenuation, the upper adjacent channel sees 16.5 dBs of attenuation, the upper alias channel sees 44 dBs of attenuation, and the upper alternate channel sees 41.7 dBs of attenuation.

Now referring to FIG 4, shown is the signal strength on the various channels in relation to the wanted signal 400 after the application of the intermediate frequency SAW filter. As is shown, the lower adjacent channel 401 and upper adjacent channel 402 are now 5 dBs stronger than the wanted signal 400. Further, the lower alternate channel 403 is now 15 dBs weaker than the wanted signal 400, while the upper alternate channel 404 is 1 dB weaker than the wanted signal. With respect to the image 405 and alias response 406 components of the signal, these have seen 44 dBs of attenuation, so 23 more decibels of attenuation are needed to suppress these to the required levels. Regarding the alternate and adjacent channels (both lower and upper), each of these signals are also required to be suppressed to –30 dBs relative to the wanted signal.

In order to avoid the second intermediate frequency SAW filter, another way of providing the required selectivity is needed in order to provide the required selectivity at the specified frequencies. Active bandpass filters are very noisy and are difficult to

implement at high IF frequencies, and applying the required LC-filter would drastically increase the number of external components. Thus, a way to avoid the application of a second intermediate frequency SAW filter is to add the requisite selectivity at the second IF.

Referring now to FIG 5, shown is a schematic diagram of one embodiment of the radio receiver portion 113. As shown in FIG. 1, the solution in the present embodiment is the application of an I/Q low IF stage and a I/Q sampling architecture. I/Q sampling means that the pair of signals (in-phase and quadrature phase components of the input signal) are sampled at four times their center frequency. One of the signals is then delayed by a period, before the signals are added together. The I/Q sampling stage provides a bit of extra filtering for both the adjacent channels and the image channel. Moreover, instead of employing the second IF SAW filter, the system uses a complex domain filter 515, which costs a fraction of what an IF SAW filter costs and can be easily integrated. Thus the second intermediate frequency SAW filter is inefficient and unnecessary, and therefore is not used.

As can be seen in FIG 5, the system assumes the availability of a super-heterodyne front end 103 (FIG. 1), which converts the 5-6 GHz signal 500 down to a first intermediate frequency of 1.5 GHz. Based upon this assumption, the first stage in the receiver applies an intermediate frequency SAW filter 505 to the input signal 500. As discussed above, the intermediate frequency SAW filter 505 provides an excellent transfer function with a narrow bandpass and steep selectivity outside the bandpass range. For example, the narrow bandpass of this filter is in the range of 30-40 MHz, while the steep selectivity of the filter is related to the narrow bandpass in that it provides for a

more precise range with regard to the bandpass characteristic of the filter. The terms narrow bandpass and steep selectivity are relative terms being defined in their relationship to most active filters having a bandpass characteristic. In this embodiment, the bandpass range must be so selective because of the close proximity of the channels. However, in other embodiments corresponding to other communication standards, the channels may not be in such close proximity and a less steep bandpass filter may provide the requisite selectivity for input into the IF sampling architecture. After the input signal has been filtered by the IF SAW filter 505, the signal is passed through an automatic gain control amplifier 510. This amplification stage 510 assures that the signal being fed into the later stages is of a relatively constant amplitude signal.

The rejection requirements outlined above can be met by analog and digital filtering in baseband. However, the problems with the image and alias channels are not addressed when the filtering is done in baseband. When converting the signal down to baseband, the modulation shifts the image channel into the wanted signal. Since the modulation causes the image channel to move into the wanted signal, the image channel should be removed as much as possible prior to baseband sampling. Assuming half of the alternate channel power passes through to the image channel, the image signal strength will be 37 dB, because half power is equivalent to –3 dB on the decibel scale. Thus, the removal of the image channel requires a selectivity of 67 dB at 30 MHz from the alternate channel. However, again, it should be understood by one skilled in the art that these decibel ratings are provided by the specifications of the HiperLAN/2 and 802.11a standards, and that the invention is not limited these standards.

In order to meet the requirements here, this embodiment converts the signal down to a second intermediate frequency prior to the second stage of filtering, as discussed above with reference to FIG 2. As indicated above, in the IF sampling architecture of the present invention, a second intermediate frequency is chosen. For this embodiment, 15 MHz has been chosen as the second intermediate frequency. At the chosen frequency, a regular active complex domain filter 515 can be applied.

In converting the signals down to this second intermediate frequency, two multipliers 520, 525 are applied, with the output of the amplifier 510 feeding one input of each of the multipliers 520, 525, and with a local oscillator 530 signal feeding the other input of each of the multipliers 520, 525. The first multiplier 520 converts the output of the amplifier 510 to an in-phase component of the input signal, while the second multiplier 525 converts the amplifier 510 output to a quadrature phase component. Next, a filter 515 is applied to add the requisite selectivity at the second IF. The complex domain filter 515 must be able to pass the wanted signal while rejecting the image and alias signals to the required degree. Restrictions in the use of complex domain filters 515 most often lie in the requisite Q factor and the maximum image rejection that can be achieved by the filter 515, which is given by I/Q balancing in the filter 515. The Q factor required for the present application, however, is quite reasonable because the channel bandwidth can be chosen at 17 MHz, with a center frequency of 15 MHz.

Mismatches prior to a local oscillator generation 530 and modulation by multipliers 520, 515 may cause leakage from the image band onto the wanted second IF signal. Once such leakage occurs, it is difficult to undo. Therefore, it is important to try to eliminate the leakage prior to it propagating through to the next modulation. However,

the importance of eliminating the leakage decreases according to the stage of the device. If the image signal is highly attenuated at the first stage of the device, there is, by definition, less signal to "leak" onto the wanted signal at the later stages.

Noting that the first intermediate frequency SAW filter 505 has already attenuated the image frequency by about 44 dBs, it becomes apparent that, by inspecting the SNR requirements, the attenuation should be in the neighborhood of 23 dBs. Further, a complex domain filter 515 can reasonably provide a rejection in the neighborhood of 25 to 30 dBs. Thus, the image channel can be reduced to the required degree by complex domain filtering 515 at the second intermediate frequency.

Referring now to FIG 6, the transfer function of the complex domain filter (a third order butterworth filter) can be seen. With regard to the transfer function of the filter shown in FIG 5, it should be noted that the center frequency is 15 MHz and the bandwidth is 17 MHz. Also shown is a leakage curve indicating the amount of power leaking onto the channel at the opposite side of the frequency axis (i.e. at –15 MHz center frequency). This leakage assumes that the in-phase and quadrature phase signals have an amplitude mismatch of +/- 2% with respect to one another.

Referring back to FIG 5, an I/Q sampling architecture 535 was employed in order to obtain some extra selectivity at the image and alias channels. As was discussed above, the I/Q sampler operates by sampling 540 the signal at 4 times the second intermediate frequency, or 60 MHz, as indicated by clock 545. A delay element 550 then delays one component of the signal for one clock period. After delaying the signal, an adder 555 then adds the two component signals back together. Finally, after adding the component signals together, the sum is quantized using an analog-to-digital converter (ADC) 560.

The ADC 560 is clocked using the same 60 MHz clock 545 as was used for the sampling 540. The I/Q sampling operation performed by circuit 535 yields a result wherein the wanted signal is amplified from the I/Q combination, while providing a notch type attenuation for the alias frequency. Ultimately, the I/Q sampler adds about 11.5 dBs of selectivity to the system over the image channel frequencies, while adding 5 dBs of attenuation to the adjacent channels.

After the signal has been digitized, it can be converted to baseband by multipliers 565, 570. The baseband signal is derived by multiplying the output of the ADC 560 by a string of coefficients. For the in-phase baseband conversion, the digital signal is multiplied by the string 0, 1, 0, -1 by multiplier 565, while the quadrature phase baseband conversion 570 is given by multiplying the digital signal by 1, 0 , -1, 0 using multiplier 570. Each of these baseband signals is then filtered by a finite impulse response filter 575, 580 to eliminate any remaining noise.

Referring now to FIG. 7, shown is an embodiment of the radio receiver of the present invention having reduced complexity. This solution sacrifices some of the performance quality for a more efficient solution in terms of cost. As can be seen in FIG. 7, the solution eliminates the delay element 550 and sampling element 540 after the complex domain filter 515 shown in FIG. 5. Instead, the in-phase and quadrature phase signals are added by adder 555 together, without the phase shift. Since the quadrature and in-phase signals are no longer efficiently combined, there is less amplification of the signal, resulting in a wanted signal that is weaker. Moreover, there is no notch effect gained, so the extra attenuation of 5 dBs at the adjacent channel and 11 dBs at the image channel is no longer a part of the system. As a result, higher order finite impulse

response filters 575, 580 are used. However, other filters provide similar characteristics may instead by used at the final stage following conversion to baseband.

Referring now to FIG 8, shown is a schematic diagram of the transceiver of the present radio architecture. With respect to the upper path of the signal, shown is the reduced complexity implementation of the receiver architecture, including the front end conversion element 820 that converts the incoming radio frequency signal to the first intermediate frequency. The signal is first received through the antenna 800 and passes through an initial bandpass filter 805. A switch is shown to represent the different input and output paths of the radio signal. On the input side, the signal passes through a low noise amplifier 810 and a second bandpass filter 815 prior to being modulated by modulator 820 by multiplying the input signal by a first local oscillator in order to convert the signal to the first intermediate frequency, 1.5 GHz. The signal then passes through the single intermediate frequency SAW filter 505 and an automatic gain control (AGC) amplifier 510 before being converted by converter 520, 525 to the second intermediate frequency comprising both quadrature and in-phase components. The conversion is performed by multiplying with multiplier 520, 525 the output of the AGC with a second local oscillator signal generator 530. At the second intermediate frequency both quadrature and in-phase components are passed through a complex domain filter 515. The outputs, in-phase and quadrature phase components, of the complex domain filter 515 are then summed by adder 555 and digitized by ADC 560. The signal is then converted to baseband by converters 565, 570 which multiply the signals by the series 0, 1, 0, -1 and 1, 0, -1, 0 to get in-phase and quadrature phase representations, respectively.

The signals are then filtered in a finite impulse response filter 575, 580 and fed into a data slicer, where the information contained within the signal is retrieved.

With respect to the lower path of the transceiver, shown is the transmitter. The transmitter is close to a mirror image of the receiver, without the more complex filtering devices. First, the signal is transferred into the transceiver in baseband. Then the baseband signal is passed through a finite impulse response filter 830, 835 before conversion by converters 840, 845 from baseband to the second intermediate frequency. The conversion by converters 840, 845 to the second intermediate frequency is performed by multiplying the in-phase and quadrature components of the signal by the series comprising 0, 1, 0, -1 and 1, 0, -1, 0, respectively. The signal is then combined and converted to analog by a digital-to-analog converter (DAC) 850 for transmission. The analog signal is then converted up by converter 855 to the first intermediate frequency by multiplication with the second local oscillator 530 signal. The signal is then amplified by amplifier 860, fed through an output SAW filter 865, and converted by converter 870 to the transmission frequency. At the transmission frequency, the signal is passed through a bandpass filter 875. The signal is finally passed through a power amplifier 880 before transmission. The power amplifier 880 can amplify to saturation in order to get the most transmission power possible out of the transmitter. On transmission, the switch is on the lower path and the signal passes through a final bandpass filter 805 before being transmitted.

It should be emphasized that the above-described embodiments of the present invention, particularly, any "preferred" embodiments, are merely possible examples of implementations, merely set forth for a clear understanding of the principles of the

invention. Many variations and modifications may be made to the above-described embodiments of the invention without departing from the scope of the invention. All such modifications and variations are intended to be included herein within the scope of this disclosure and the present invention.